United States
Naval Postgraduate School

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AN EIGHT-CHANNEL SAMPLED-DATA ACOUSTIC TELEMETRY SYSTEM
FOR
DEEP-WATER BIOMEDICAL AND ENVIRONMENTAL APPLICATIONS

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Leslie James Reading
Lieutenant Junior Grade, United States Navy
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ABSTRACT

A general-purpose acoustic telemetry system is presented. The carrier of one hundred forty-five kilohertz is frequency shift keyed at an average rate of once a millisecond, the subcarrier being pulse-width modulated by seven channels of analog data and one synchronization channel. The receiver is automatically synchronized and various error detecting schemes are employed.

The system is designed to be battery operated at the transmitter and is specifically intended to telemeter physiological and environmental data from a free swimming diver to the surface. The theoretical lateral design range is three hundred meters.

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I. INTRODUCTION

In recent years the diving of Navy divers to ever increasing depths has resulted in a tightening of safety precautions in the face of the many possible dangers associated with a high-stress, hyperbaric environment. One safety precaution is real-time physiological monitoring so that personnel not subject to the high stresses can oversee the diver's progress. The diver can then be warned to return to safety in the event that complications occur of which the diver may not be immediately aware. In this respect an underwater telemetry system is a valuable safety device as well as being a device that can yield significant medical and scientific data concerning the nature and degree of stress that the diver is undergoing.

Two significant applications of underwater acoustic telemetry have been made in the past, one an fm/fm telemetry system experimented with by Bendix (Ref. 1) and the other a very practical sampled-data system for ocean bottom seismometer readings (Ref. 2). Beyond this very little work has been published as a result of the limited financial returns of research in this area compared to the large financial outlay necessary to get meaningful data.

The system described in this work is a general-purpose sampled-data telemetry system intended for but not specifically limited to the telemetry of physiological and enviromental data. It is designed to be used in the following practical situations:

1. a free-swimming diver in either deep or shallow water with an acoustic data link to the surface

2. a diver tethered with an umbilical cord in very deep water with either an acoustic data link to the surface or a hard-wire data link through the PTC (personnel transfer capsule) to the surface

3. a diver tethered to or with an electrical connection to an underwater vehicle equipped with a high-fidelity tape recorder with the intention of reproducing the monitored data at a later time.

In all these situations the use of the system includes the recording of the monitored data with the capability of reproducing that data at a later time.

The practical realization of the system led to a continuous revision of the specifications leading to the following guidelines:

1. design range, 300 meters lateral

2. eight channels, time-division multiplexed with one of the eight channels devoted to a synchronization channel

3. battery-operated transmitter with a battery lifetime of two hours.

With the above guidelines and with the practical limitations of available material, a system was designed, built, and tested and is presently being optimized.

II. SYSTEM PARAMETERS

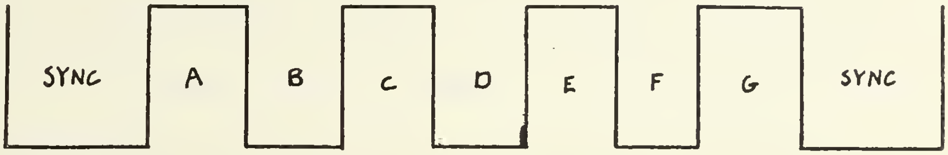
A. BANDWIDTH

The acoustic carrier is frequency shift keyed at a rate determined by the subcarrier which is in turn a form of pulse-width modulation. The amount of frequency shift used is 10 KHz centered about a carrier frequency of 145 KHz, this center frequency being determined initially by the characteristics of underwater ambient noise and finally by the transmitting and receiving response of the available hydrophone and transducer. The transmitting transducer is operated below resonance at a loss in efficiency but at a point where the transducer's impedance characteristics are quite insensitive to changes in frequency. Minimum distortion is achieved as a result of this bandpass characteristic. The receiving hydrophone response is essentially flat from 85 KHz to 180 KHz which tends to minimize any additional distortion at the receiver.

High resolution in transmitted data is required, with the capability of handling electrocardiagram signals being the limiting factor in maximum bandwidth. The number of channels is limited to eight as a result of the distribution scheme used in which a binary register drives an analog switch. This analog switch is an integrated circuit and is available in no more than eight channels. A lesser number of channels would have resulted in inefficient use of the capabilities of the circuit.

The subcarrier is pulse-width modulated such that the analog information is contained in the time interval between transitions, yielding efficient use of the available system bandwidth.

This is done by comparing the analog voltage with a ramp generator and resetting the ramp generator when the signals are of the same value. The reset pulses are used to clock a flip-flop, the output of which constitutes the subcarrier.



As is shown in the illustration, the channel devoted to synchronization is longer than the data channels. This is true in all cases since a data channel is limited to an absolute maximum value and this maximum is always less than the width of the synchronization channel. On the other hand, there is also an absolute minimum value that a data channel can obtain such that all channels can be accounted for at the receiver.

Under the simplification that the average time interval between transitions is one millisecond, this value being sufficient to transmit electrocardiagram data (acceptable to the personnel for whom the system is intended,) the fundamental frequency of the subcarrier is 500KHz. As an approximation using linear frequency-modulation theory the modulation index would be as follows:

$$\beta = \text{modulation index} = \Delta f/f_m = 10 \text{ KHz}/500 = 20.$$

For this large a value of β the approximation holds that the bandwidth is approximately equal to twice Δf or 20 KHz, thereby determining the bandpass response of the filter at the input of the receiver. Tests with the system have shown that the distortion of the received signal as a result of this bandpass filter is quite small.

B. ENVIRONMENT

Underwater ambient noise in the vicinity of the carrier frequency (145 KHz) is essentially thermal and is expressed as follows:

$N_o = -115 + 20 \log (f)$ - f in KHz, No in μbar/Hz.

Using the given system bandwidth of 20 KHz and carrier frequency of 145 KHz, the ambient noise referred to the input of the receiver can be calculated as follows:

$$N_o = -115 + 20 \log (f) = -115 + 20 \log (145)$$

$$= -72 \text{ db } \mu\text{bar/hz}$$

$$N = N_o + 10 \log (BW) = -72 + 10 \log (2 \times 10^4)$$

$$= -72 + 40 + 3 = -29 \text{ db } \mu\text{bar.}$$

Conventional frequency-modulation theory requires a fifteen-decibel signal-to-noise margin at the receiver for good reception.

Therefore, at the receiver the signal level (S) is required to be;

$$S = -29 + 15 = -14 \text{ db } \mu\text{bar.}$$

The absorption coefficient of sea water at these frequencies is given as follows:

$$\alpha = 7 \times 10^{-2} \text{ db/meter.}$$

The attenuation for a range R is then

$$N_W = 20 \log(R) + \alpha(R),$$

where $R = 300$ meters $= 3 \times 10^2$ meters.

$$N_W = 40 + 20(.475) + 7\times10^{-2}(3\times10^2) = 70.5 \text{ db.}$$

The required intensity at the transmitter is

$$I = -14 + 70.5 + 56.5 \text{ db } \mu\text{bar.}$$

From manufacturer's data, the minimum transmitting response is 40-db μbar/volt. Therefore, the voltage required to drive the transducer is +16.5 db volt.

$$+16.5 = 20 \log (V)$$

$$V = 6.7 \text{ volts} = 19 \text{ volts (p-p)}$$

The peak-to-peak voltage is of particular importance since it is the limiting factor in the transmitting link when no transformer is used to match to the the transducer.

The receiving hydrophone response is -110 db volts/μbar as specified by the manufacturer. Therefore the voltage available at the hydrophone terminals at the maximum range is -124 db volt or about one-half microvolt. This requires the use of a low-noise preamplifier with the hydrophone.

The system parameters are calculated for the specific case of omnidirectional hydrophones since this is the case for the normal operating mode. The parameters are tabulated as follows:

Carrier frequency 145 KHz

Voltage applied to transducer 6.7 volts

Voltage at receiving hydrophone for the maximum range 0.63 microvolts

System bandwidth 20 KHz

Design range 300 meters

III. GENERAL FUNCTIONAL DESCRIPTION

A. TRANSMITTER

Figure one constitutes the flowchart for the transmitter and includes all the functions of the transmitter package with the exception of the voltage regulator. A regulator is necessary to insure that the entire unit is supplied with a stable, well regulated voltage in the face of changing battery voltages or other supply voltages which are incompatible with the circuitry. This regulator is a high-gain feedback, series type regulator which dissipates little power in the process of regulation and supplies a virtually constant thirty volts to the circuitry from a thirty-two to forty-five volt supply, above which value the dissipation of the series regulator becomes a serious limitation.

As a result of the fact that the system is intended to be a general-purpose telemetry system, certain forms of peripheral circuitry are necessary in order to adapt the system to any number of practical situations. The most notable of these is that in which a NASA Gemini-series electrocardiagram monitor is used to detect biopotentials on the diver. This device is available in large numbers and is currently being used by the personnel for whom the system is intended. This "signal conditioner" is an effective impedance-transformation device but yields useful signals of only sixty millivolts at its output, which makes necessary an additional interface unit to amplify the signal. More amplitude is required to fully utilize the capabilities of the system, specifically to raise the signal level to a value somewhat above the inherent noise level at the receiver.

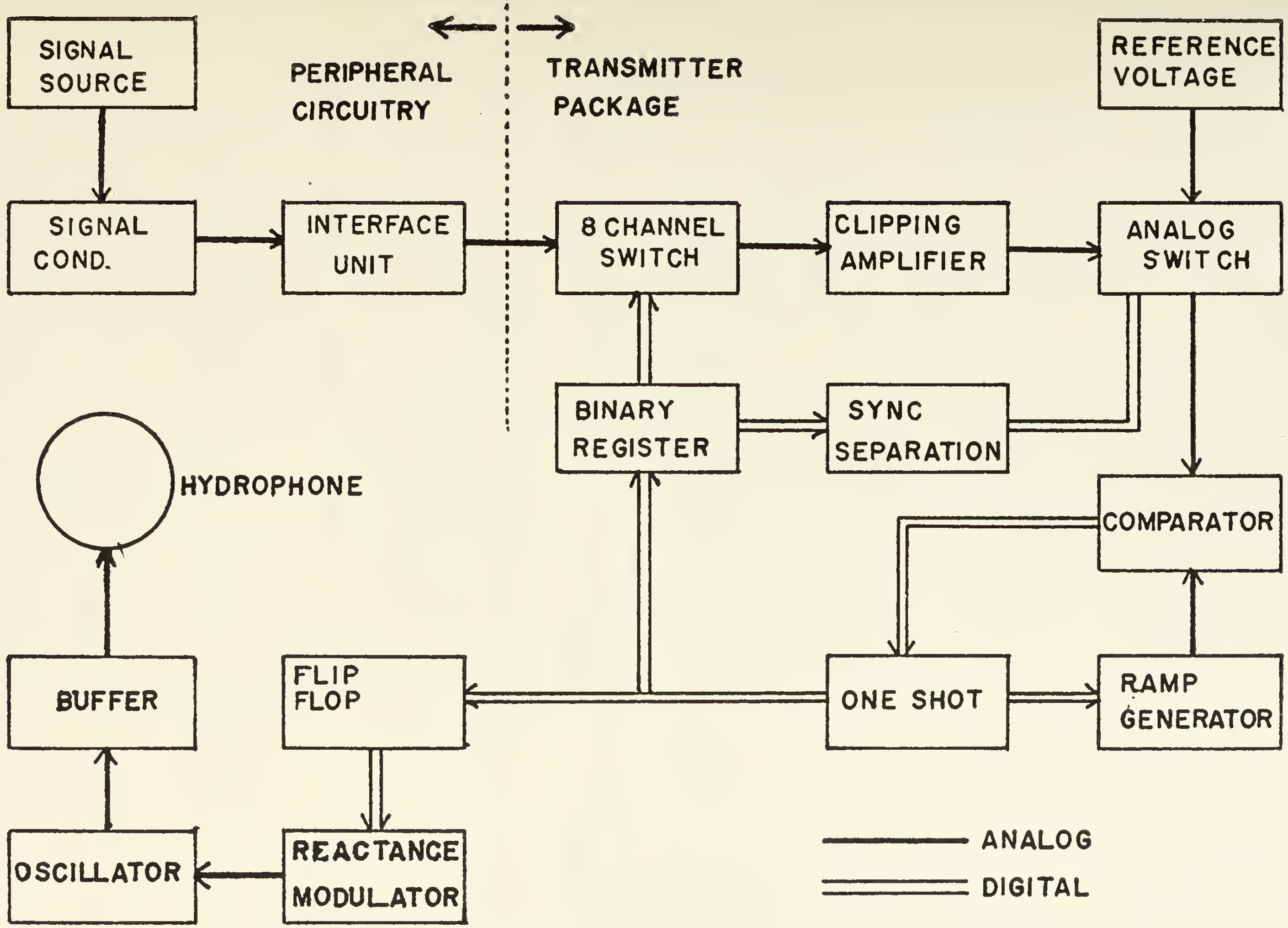


FIGURE ONE

TRANSMITTER FUNCTIONAL DIAGRAM

The first function which the transmitter performs is to sequentially sample the seven data inputs applied to the eight-channel analog switch, the eighth channel being assigned as the synchronization channel. This is done with a single integrated circuit which is described in appendix C. The only other input required by this integrated circuit is the binary address of the particular channel to be sampled, the binary decoding taking place entirely within the device resulting in significant savings in circuit complexity.

After sequential sampling, the signal is processed by a clipping amplifier which insures that the output voltage is always less than that voltage which corresponds to a synchronization channel, and is always somewhat greater than the minimum voltage of the ramp generator. This results in a unique synchronization pulse width, greater than the data-channel pulse widths. It also specifies that there is an absolute minimum pulse width such that in the receiver, all channels are accounted for when the channels are distributed according to their numerical order and there is a minimum probability of feedthrough from one channel to another at the receiver. Additionally, a malfunction in a sensing device which would normally overload the transmitter would only disenable one channel and not the entire unit.

The nature of the clipping amplifier specifies the two general transmitter packages which are available. The first package developed and tested used a clipping amplifier which was non-inverting and had a gain of ten, such that the transmitter had a very high input impedance and was capable of interfacing directly with NASA Apollo biopotential

monitoring devices which have useful output voltages of about one volt peak-to-peak. The other package available uses an inverting configuration with unity gain and low input impedance and any additional gain which may be desired can be provided for with peripheral circuitry as is the case with the NASA Gemini biopotential monitoring devices.

An additional analog switch is required in order to interpose the automatic synchronization feature with minimum circuitry. This is a discrete, two-channel analog switch which passes the output of the clipping amplifier to the comparator in all cases except when the synchronization channel address occurs in the binary register at which time it passes a reference voltage which is slightly greater than the maximum possible voltage from the clipping amplifier.

The comparator, one shot, and ramp generator constitute the critical portion of the unit which clocks the binary register and yields an output to a flip-flop which forms the subcarrier. The one shot is necessary in order to allow sufficient reset time for the ramp generator to return to its full reset value since the output of the comparator is a relatively small amount of current of short duration. It is in this circuitry that the accuracy and linearity of the system lies and additional modifications are necessary in the future in regards to this portion of the transmitter in order to reduce the complexity, size, and current requirements. In its present configuration an operational amplifier is used as a ramp generator with an analog switch in the feedback path to reset what is essentially an integrator with a constant input.

The output of the comparator clocks both the binary register and an additional flip-flop which is available by virtue of the fact that dual J-K flip-flops are used in integrated-circuit form. The flip-flop not used in the binary register can be independently loaded to drive the reactance modulator without affecting the analog switch, principally because there is an incompatibility in the voltage levels necessary to drive the reactance modulator and the analog switch.

The oscillator is a tuned L-C oscillator whose resonant frequency varies according to the voltage applied to the reactance modulator. The oscillator then drives a switching circuit which supplies square-wave input power to the transducer, a switching circuit being used in order to provide minimum thermal dissipation in the buffer.

The regulated thirty volts is too large to supply the oscillator since a reactance oscillator using this value of voltage would encounter overshoots of sixty volts, above the breakdown voltage of the available economy transistors. For this reason a voltage step-down circuit is necessary although it is not shown in figure one.

B. RECEIVER

1. General

The complexity which is saved in the transmitter appears as a necessary complexity in the receiver to insure reliable signal reception and processing. For the most part the circuitry in the receiver as well as in portions of the transmitter is overdone, primarily as a result of the fact that the design was done without a clear idea of available signal waveforms and amplitudes. Each portion of the receiver was designed as a general-purpose device which did not require strict specifications.

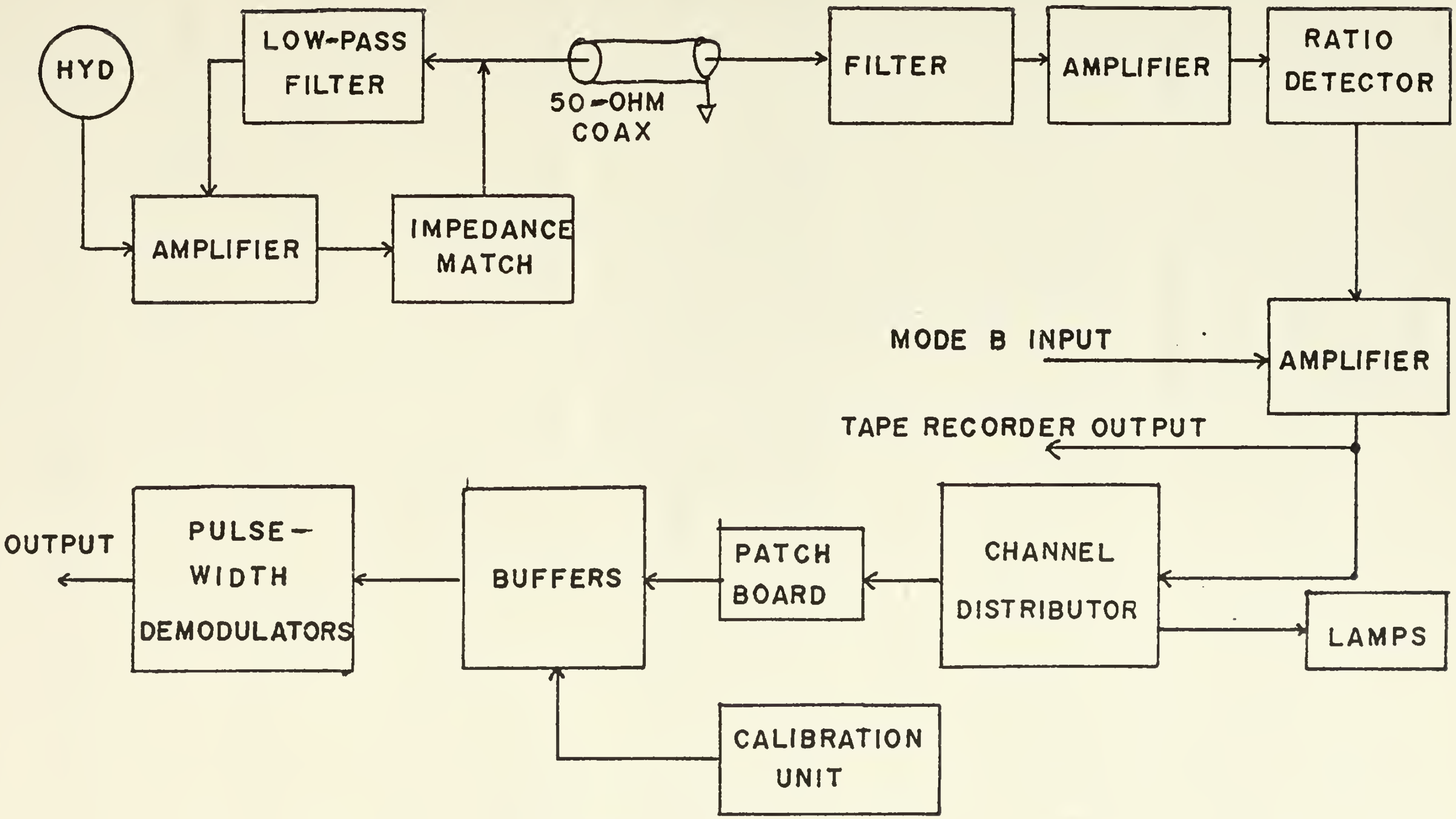


FIGURE TWO

RECEIVER FUNCTIONAL DIAGRAM

For a 15-db signal-to-noise margin at the receiver, a signal level of -124 db volt is available for processing. At such small signal levels a low-noise front end is absolutely necessary. A preamplifier with considerable gain lends itself well to this effect and in addition impedance matches to the coaxial cable connecting the receiver to the preamplifier.

Figure two illustrates the overall layout of the receiver while avoiding the complexity of the channel-distribution and error-check and pulse-width demodulator circuits which will be covered in more detail in a later section.

In regards to the preamplifier, it should be noted that it is designed to operate without using batteries at the hydrophone but rather by filtering a bias potential from the coaxial cable for power while feeding the high-frequency output on the cable. The degree of amplification is such that the final stages operate in saturation even for minimal signal levels in order to provide a constant-amplitude signal to the ratio detector. The output of the ratio detector involves frequency modulation only and not amplitude variations.

The magnitude of the output of the ratio detector is small and requires sizeable amplification prior to processing. The subcarrier, which is the output of the ratio detector, is amplified to saturation in order to provide definite transition times. The information is essentially contained in the time interval between transitions and high resolution depends on determining these transitions reliably and consistently.

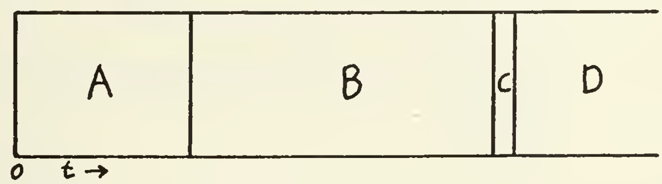
The channel-distribution and error-check unit has eight outputs including the synchronization channel, but a slight amount of feedthrough

in the analog switch makes necessary slight filtering in order to reduce the high-frequency noise in the signal applied to the pulse-width demodulators. This is accomplished by the buffers which also provide a convenient place to include a programming matrix and a calibration unit feature.

There are two general modes of operation; that in which the input is a frequency shift keyed carrier, and that in which the subcarrier is applied directly. Mode B is the case in which the subcarrier is already available, whether from hard-wire telemetry or from a previous tape recording, and an appropriate input location is supplied.

2. Channel Distribution and Error Check

The channel-distribution and error-check unit operates through the use of "windows". A ramp generator and comparators are used to determine the suitability of the input data. The previous transition time is used as a reference and the immediately following transition time is interpreted according to the following illustration:



A. High-frequency noise

- reject data and await synchronization channel prior to proceeding

B. Acceptable data

C. Synchronization channel

- reset binary register and proceed

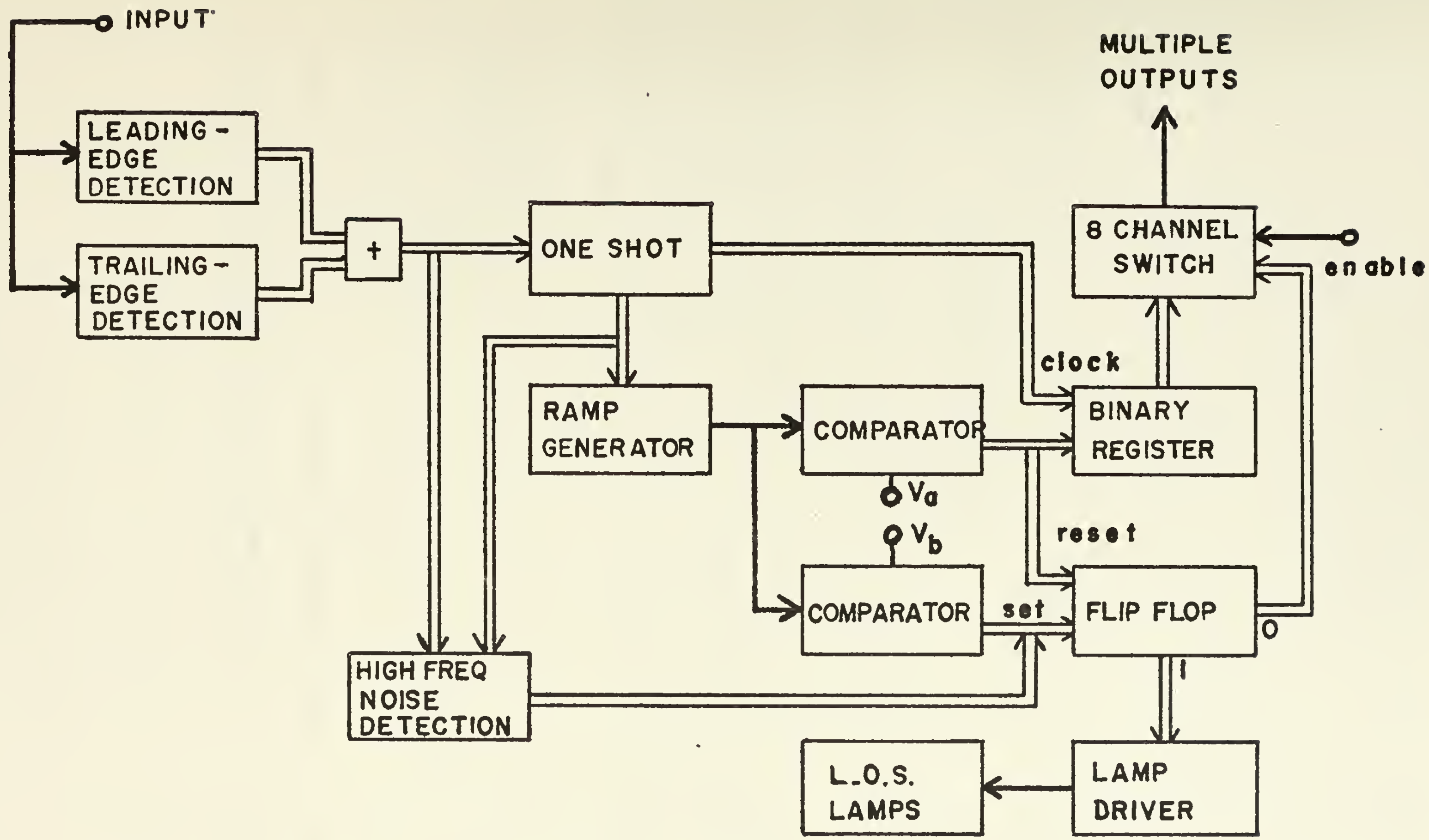


FIGURE THREE

CHANNEL DISTRIBUTION AND ERROR CHECK

FUNCTIONAL DIAGRAM

D. Low-frequency noise

- cease processing data and await synchronization channel prior to proceeding

If unacceptable data is detected the entire unit is disenabled and a synchronization channel must be received prior to resetting the unit and proceeding from the beginning. When the unit has been disenabled, a pair of lamps on the receiver light to indicate an interruption in data flow.

3. Pulse-Width Demodulator

The pulse-width demodulator is a circuit which features high resolution and is capable of reliable operation in the face of intermittent input pulses or input pulses of other than a constant duty cycle. The unit uses a ramp generator, the voltage of which is maintained by a sample-and-hold circuit at the end of the data pulse, until the end of the following data pulse. A high-quality buffer is also provided to avoid loading of the sample-and-hold circuit and to ease interfacing with any low-input impedance recording equipment.

The output of the pulse-width demodulator is plus or minus five volts maximum for reliable input data, the same value as the input for the Gemini interfaced transmitter and ten times the input value for the Apollo matched transmitter.

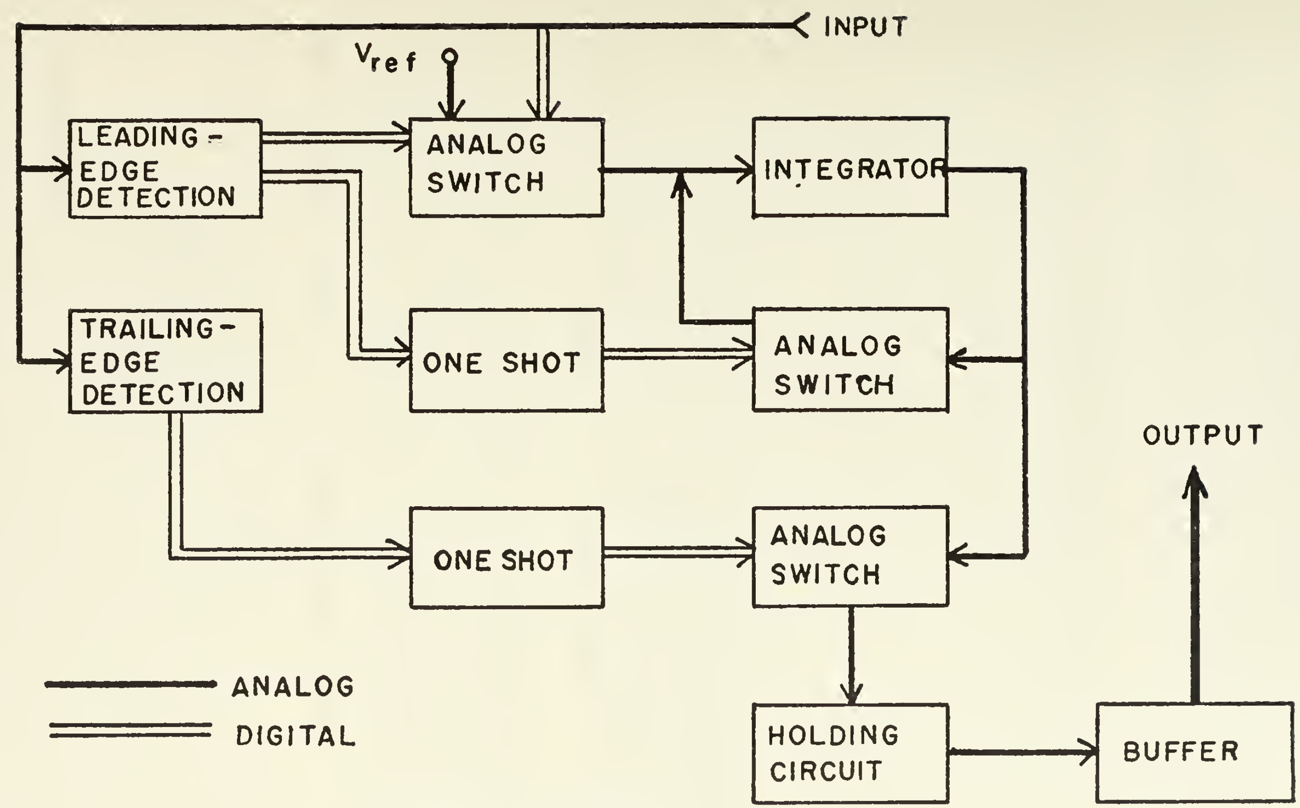


FIGURE FOUR

PULSE-WIDTH DEMODULATOR FUNCTIONAL DIAGRAM

IV CIRCUIT DESCRIPTION

A. TRANSMITTER

1. Voltage Regulator, Schematic A-1

The regulated voltage is determined by the zener diode Z which breaks down at its reference voltage and reduces the current in the base of the series regulator T1. Capacitiors C1 and C2 act not only to provide a low AC impedance for the load but also to reduce the noise from the zener diode which would normally become greatly amplified. R3 serves to protect the zener diode in the event of a malfunction and automatic current limiting is provided by R1 since the largest amount of current that can pass occurs when T2 is in saturation.

The charge rate of the regulator was adjusted such that the rise in voltage upon sudden application of a battery potential is somewhat greater than the ramp generator. If this were not so, the comparator would not trigger the one-shot and reset the integrator which is required in order to cycle to the next channel. For this reason, when a power supply is used in lieu of batteries for test purposes, the voltage on the power supply must be adjusted first and then connected to the battery terminals or the system will not recycle.

2. Clipping Amplifier, Schematic A-2

The clipping amplifier was designed around a Fairchild μA741 operational amplifier (see appendix B) and the schematic shows the configuration for the Apollo matched transmitter. The non-inverting configuration is the least desireable for a clipping amplifier since voltage overloads can destroy the integrated circuit. This is not the case for the inverting configuration where the amplification is by

virtue of current, not voltage. While the savings in component requirements are not great, the increase in reliability and the decreased zener diode tolerances make the inverting case preferable.

3. Ramp Generator, Schematic A-3

The μA741 operational amplifier acts as an integrator with a constant input determined by Z2 and a reset voltage set by Z1. It was intended that Z1 and Z2 in this circuit and Z3 in schematic A-2 track each other with temperature but this has not necessarily been the case.

Transistor T1 constitutes a simple analog switch which resets the integrator when the gate of T1 is switched positive by the one shot. The pinch-off voltage for the field-effect transistor must be less than the zener voltage established by Z1, less an additional 0.6 volts for the diode connected in series with the gate of T1. The R_{ds}(on) of about 200 ohms for T1 limits the reset time which requires that the pulse width from the one shot be of a reasonable value.

4. Comparator and One-Shot, Schematic A-3

A reset time of about 0.2 milliseconds was determined to be sufficient to reset the integrator to its full reset value. This time period is established by C2 and R5 and the subsequent diode is to prevent the destruction of T4 by excessive breakdown from the emitter to the base. The voltage drops to minus thirty volts with respect to the emitter and with a typical breakdown potential of five volts, it is necessary that the diode be in the current path. The diode cannot be in the emitter because sixty volts would be seen from the collector to the base during the initial transition and this is well over the collector breakdown rating of forty volts.

The one-shot is initially triggered by a current pulse from T2. This can only occur when the emitter of T2 is positive with respect to the base and with typical differential voltages of twenty-five volts from the base to the emitter, a diode is required to protect the transistor.

5. <u>MOS Integrated Circuits</u>, Scematic S-4

MOS integrated circuits were chosen because of their high functional density and low power requirements. Being thirty-volt devices, they interface well with thirty-volt operational amplifiers. High speed is not a critical requirement in this application and although the 3705 and the 3101 were not designed to be interfaced as a result of supply voltage incompatibilities, they do so nicely if one exceeds their recommended ratings slightly. (See appendices C and D.)

6. <u>FSK Oscillator</u>, Schematic A-5

The voltage levels from the MOS integrated circuits do not cover the full dynamic range of transistors and the zener diode is present to insure that T1 will "cut-off" when the JK flip-flop is in the "off" state even though the corresponding voltage is not quite V+. For the same reason, zener diodes are used in the three input NAND gate in schematic A-2 which serves to implement the synchronization function.

The noteworthy item about this circuit is the use of a field-effect transistor, T4, to change the frequency of oscillation. The reactance of C3 is much greater in magnitude than the resistance of R9 causing the gate to be driven ninety degrees out of phase, which makes the reactance modulator look like an inductor. As a result of the fact

that the field-effect transistor is a square-law device, the transconductance can be varied by changing the gate-to-source bias potential which in turn changes the value of reactance reflected to the tuned load of T3. Of all the techniques tried, this method has shown itself to distort the waveform the least when switched.

Transistors T7 and T8 are capable of driving the transducer from a switching voltage but as a result of the fact that their bases are directly connected, they will of themselves dissipate no power unless they are driving a load. This push-pull arrangement has worked quite well for driving heavily capacitive loads such as a ceramic transducer.

B. RECEIVER

1. Hydrophone Preamplifier, Schematic A-6

As a result of the fact that the preamplifier must be capable of handling very small signals (0.63 micro-volts) a field-effect transistor front end is required. Additionally, R3 and R10 are low-noise wire-wound resistors. This combination results in a very low-noise circuit. High input impedance is not required as can be observed from the impedance characteristics of the hydrophone as shown in figure six.

The coaxial cable has a fifty-ohm characteristic impedance and is matched at both ends to this impedance. This is done at the preamplifier by using a buffer, T4. The output is fed ahead of the low-pass filter such that there is minimal feedback of the signal while the DC bias on the cable is available to power the preamplifier.

2. <u>Four-pole Filter and Amplifier</u>, Schematic A-7

By using the image parameter method, a very effective input filter was achieved. The variable inductors can be adjusted such that the signal on the drain of T1 shows no sign of distortion, indicating the validity of the bandwidth approximation.

C11 exists to eliminate parasitic oscillations which occur from feedback from T4 to T3. R7 serves not only to filter parasitics from the following stages but also to serve as a current-limiting device in the event that the input is short circuited.

3. <u>Ratio Detector</u>, Schematic A-8

The ratio detector is a tuned-primary, untuned-secondary device such that it will operate effectively in the face of possible drifts in the transmitter frequency. Transistors T2 and T3 drive the ratio detector to its full capabilities but subsequent amplification of the subcarrier is still required.

4. <u>Data Distributor</u>, Schematic A-9

This circuit through T6 constitutes a leading and trailing-edge detector with a one-shot that triggers at each subcarrier transition.

As a result of the fact that the one-shot triggers on the leading edge of the triggering pulses, high-frequency noise detection must have a sufficient delay to allow the initial pulses to return to zero volts before functioning. This is accomplished by limiting the output of transistor T7 with a zener diode so that the voltage of C5 is not sufficient to yield an output if the gate is enabled until the initial pulse has had time to return to zero volts.

5. Data Distributor Ramp Generator, Schematic A-10

The "windows" that are used to determine the acceptability of the input data are established by transistors T2 and T3. Essentially, the delay of the one-shot establishes the first "window" indicating high-frequency noise and the voltage level set in T2 indicates low-frequency noise. If T3 triggers and T2 does not, then the received pulse corresponded to a synchronization channel.

6. MOS Integrated Circuits, Receiver, Schematic A-11

If an eight-channel analog switch can have eight inputs and one output, it also follows that it can have one input and eight outputs. The advantage here is not so much the circuit complexity as it is the the fact that the 3705 and the 3101 can be made to interface directly and the 3705 also has an all-channel blanking capability that permits the device to allow no outputs at all when the output enable is in the "low" state.

7. Buffer Amplifiers, Schematic A-12

The buffers provide clean waveforms to the pulse-width demodulators without interference between channels. An additional feature is that as many as four non-consecutive channels can be added together at the transmitter and receiver in order to improve the frequency response of the system for a given set of data if needed. The minus fifteen volts applied to R1 comes from the calibration unit so that by applying a square wave at this point, all channels will receive the same pulse width and they can be calibrated accordingly.

8. Signal Conditioner, Scematic A-13

This circuit is a leading-and trailing-edge detector as well as a device which provides a ramp generator "hold" when the sampling takes place. The outputs drive the one-shots and the reset integrator as well as establish reference voltages for the integrator.

9. Ramp Generator, Schematic A-14

The ramp generator uses an analog switch to reset the integrator in which case the pinch-off voltage of transistor T5 must be less than the zener voltage established above the minus fifteen-volt supply by Z1 in schematic A-13. For the same reason the pinch-off voltage of T2 must be less than the voltage established by Z2, both in schematic A-13, this being the requirement to implement a hold for the integrator in the ramp generator.

10. Sample and Hold, Schematic A-15

When T5 is pinched off, the leakage currents approximate the reverse bias currents in T6 and the effect is to tend to cancel any voltage drift by current going into or leaving C5 which should itself be a low-leakage capacitor. The constant-current source, T7, greatly increases the dynamic range of the sample-and-hold circuit and at the same time guarantees a constant offset voltage through T6 which insures unity gain through the sample-and-hold unit.

11. Calibration Unit, Schematic A-16

Three pulse widths are available from the calibration unit in order to give a three-point determination of the characteristics of the pulse-width demodulator. With this data, a sequential approximation can be made by adjusting the ramp-generator delay and the ramp-generator rate to reproduce input data with high accuracy.

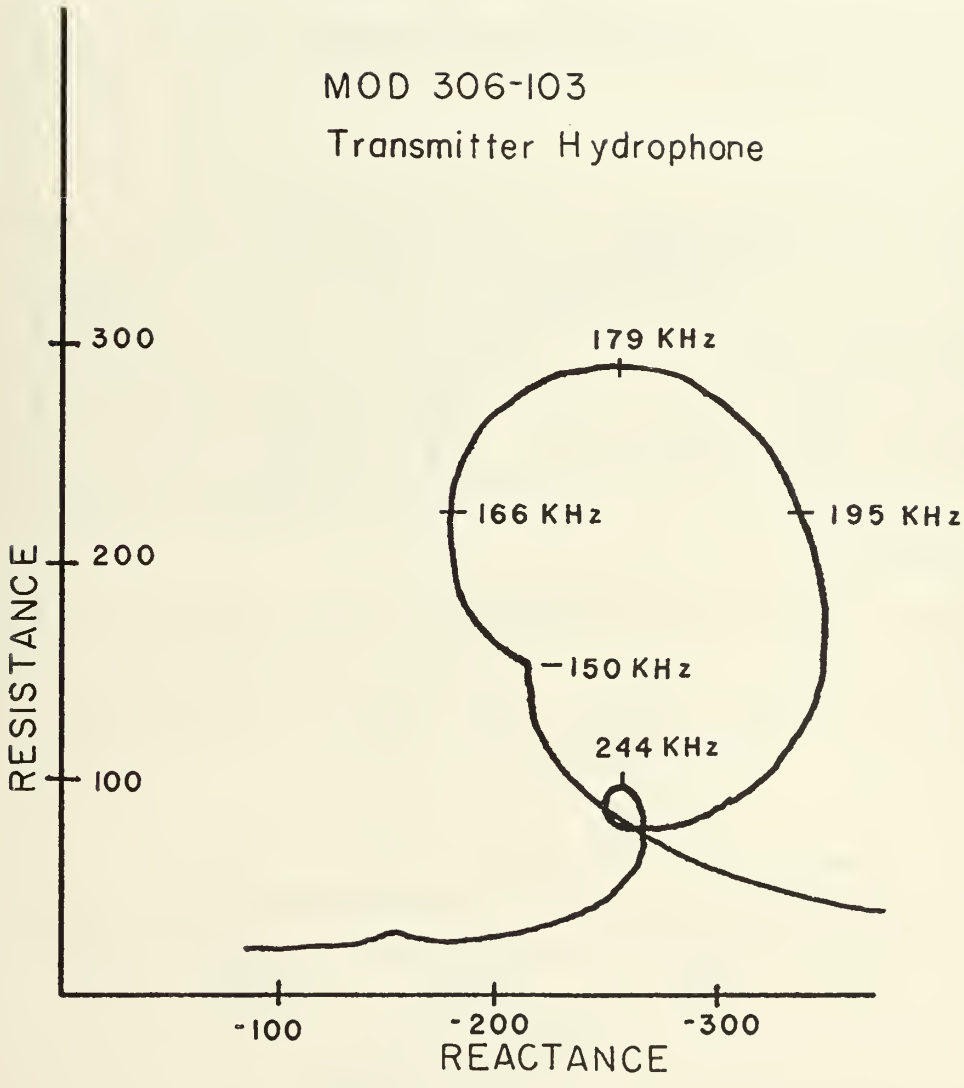


FIGURE FIVE

TRANSDUCER IMPEDANCE CHARACTERISTICS

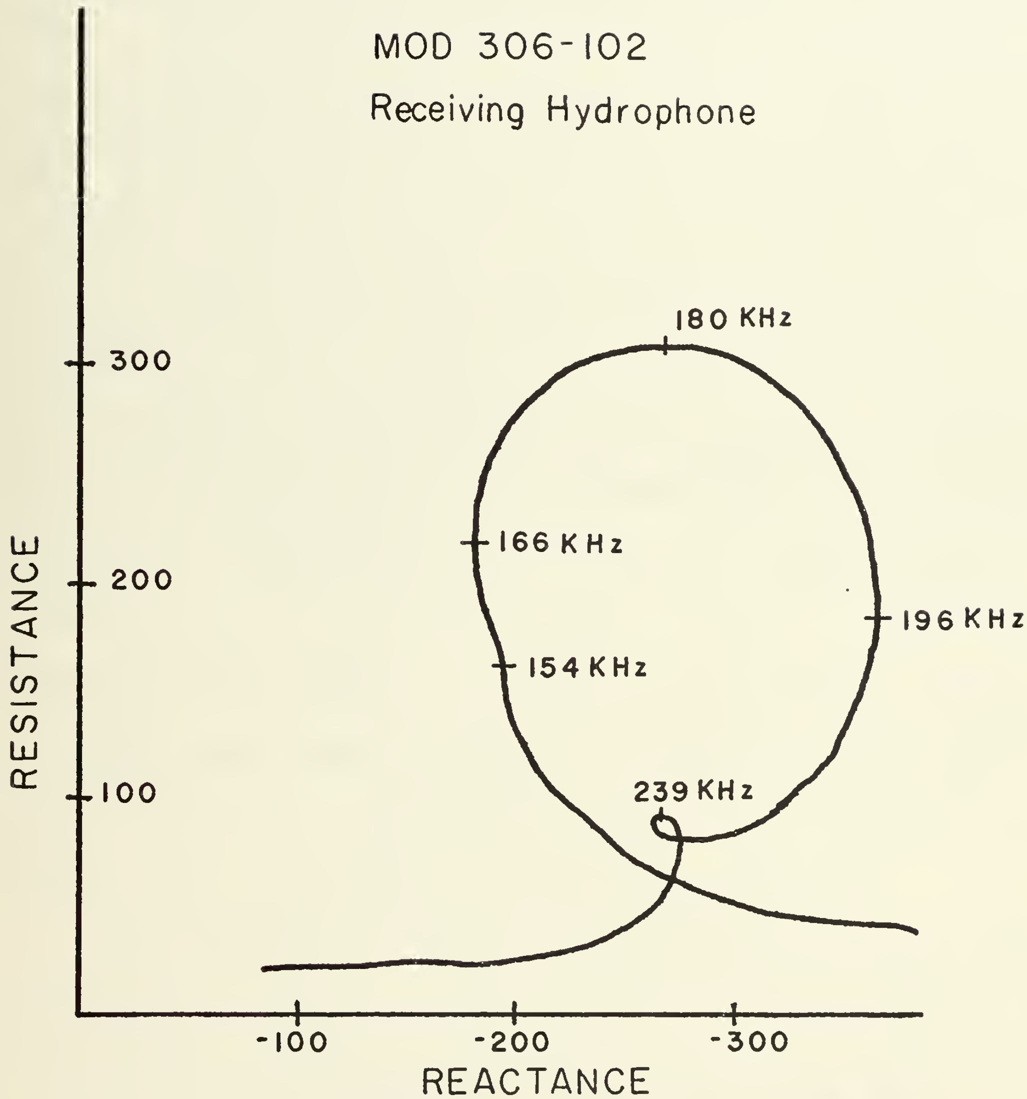


FIGURE SIX

RECEIVER IMPEDANCE CHARACTERISTICS

V EXPERIMENTAL RESULTS

Experimentation was performed with a Gemini matched transmitter, with an electrocardiagram signal generator providing the input waveforms. Two operational amplifiers were used to bring the voltages from the signal generator up to a useful value that was compatible with the transmitter input. A single 42-volt mercury battery was used as the power supply for the transmitter with the exception of the transducer buffer which was supplied with power from two nine-volt batteries connected in series. The entire package was sealed in a stainless-steel cannister four inches in diameter and eight inches deep. The RG-74 cable to the transducer was passed through the casing with a teflon stuffing gland. The lid to the cannister was provided with an "O" ring seal. The specification given when the cannister was machined was that it be water tight to pressures corresponding to depths as great as a thousand feet, and the transmitter has been tested to depths of as much as four hundred feet with no sign of leakage.

The preamplifier was sealed in an additional cannister smaller in size but with equal water-tight integrity. This cannister was attached to the end of two hundred feet of RG-58 coaxial cable.

The first test was performed in shallow water with a depth of about twenty feet, the transmitter and receiver both being at a depth of about ten feet. Commencing with a range of eight feet, the transmitter was gradually moved away while the subcarrier was displayed on an oscilloscope and recorded on a tape recorder. The quality of the signal ranged from excellent to very poor with a maximum effective range of about thirty feet. Multipath transmission effects were observed,

while monitoring the signal received from the preamplifier, which were manifested by large amplitude variations in this signal.

The transducer and hydrophone were then placed on the bottom (sand) in order to rely upon a single reflected signal from the surface boundary. The quality of the signal received at the input of the receiver was greatly improved although the effective range was not changed, but it should be pointed out that the surface of the water was quite smooth. An analysis of the front end of the receiver under these conditions indicated that parasitic oscillations were primarily responsible for the short range since the signal at the input of the receiver had to be sufficiently large to swamp out the oscillatory nature of the amplifiers.

A new preamplifier and front end were installed prior to making deep water tests and the new preamplifier featured impedance matching to the cable and tuned input stages. When used, however, the monitored subcarrier yielded an audio output that, when listened to on the tape recorder, was the subcarrier with superimposed music from a local radio station, KMBY. This occurred with the hydrophone at a depth of about fifty feet and at a distance from the radio station of two miles.

A new preamplifier was then developed which was broad banded and a new front end to the receiver featuring a four-pole passive filter was designed. Tests were run in Monterey Harbor in water depths of twenty to thirty feet.

Audio monitoring of the subcarrier proved to be an excellent means of determining whether the transmitter had been placed in the water and the presence of a signal in the water was noted audially at ranges up to one hundred fifty meters. The small size of Monterey

Harbor ruled out ranges farther than this as well as the fact that multipath reflection effects were so severe that they made the subcarrier ineffective at virtually all ranges. Operation, even at close ranges, was not possible as a result of the reflections from the hull of the vessel from which the receiver was suspended.

A deep water test was then made from a sixty-three foot vessel with the intent of operating the system in an environment free from multipath reflections. The wave height was three to four feet with occasional white-caps and on a scale comparable to the acoustic wavelength of the carrier the surface was quite turbulent.

The initial separation of the transmitter and receiver was twenty feet and severe multipath reflection was observed at a depth of three feet for both. The receiver was then lowered to one hundred seventy feet and the transmitter was lowered in fifty foot increments to four hundred feet. At no time was the subcarrier effectively demodulated although the presence of the signal was obvious. Analysis of available waveforms resulted in the conclusion that the unacceptable subcarrier was a result of a loss of signal from the preamplifier. This could be the result of excessive drain on the low-pass filter from the preamplifier being over-driven, which would cause the bias on the amplifier stages to depart from class "A" bias conditions to class "C". The possibility of second-order effects is also present such that a limit cycle might also occur as a result of overdriving the preamplifier.

The possibility of phase cancellation as a result of anomalies must also be considered in light of the sound-velocity profile taken during the deep water test. For two paths of identical lengths, a

difference in velocity of only 0.75 meters per second is necessary to cause complete phase cancellation.

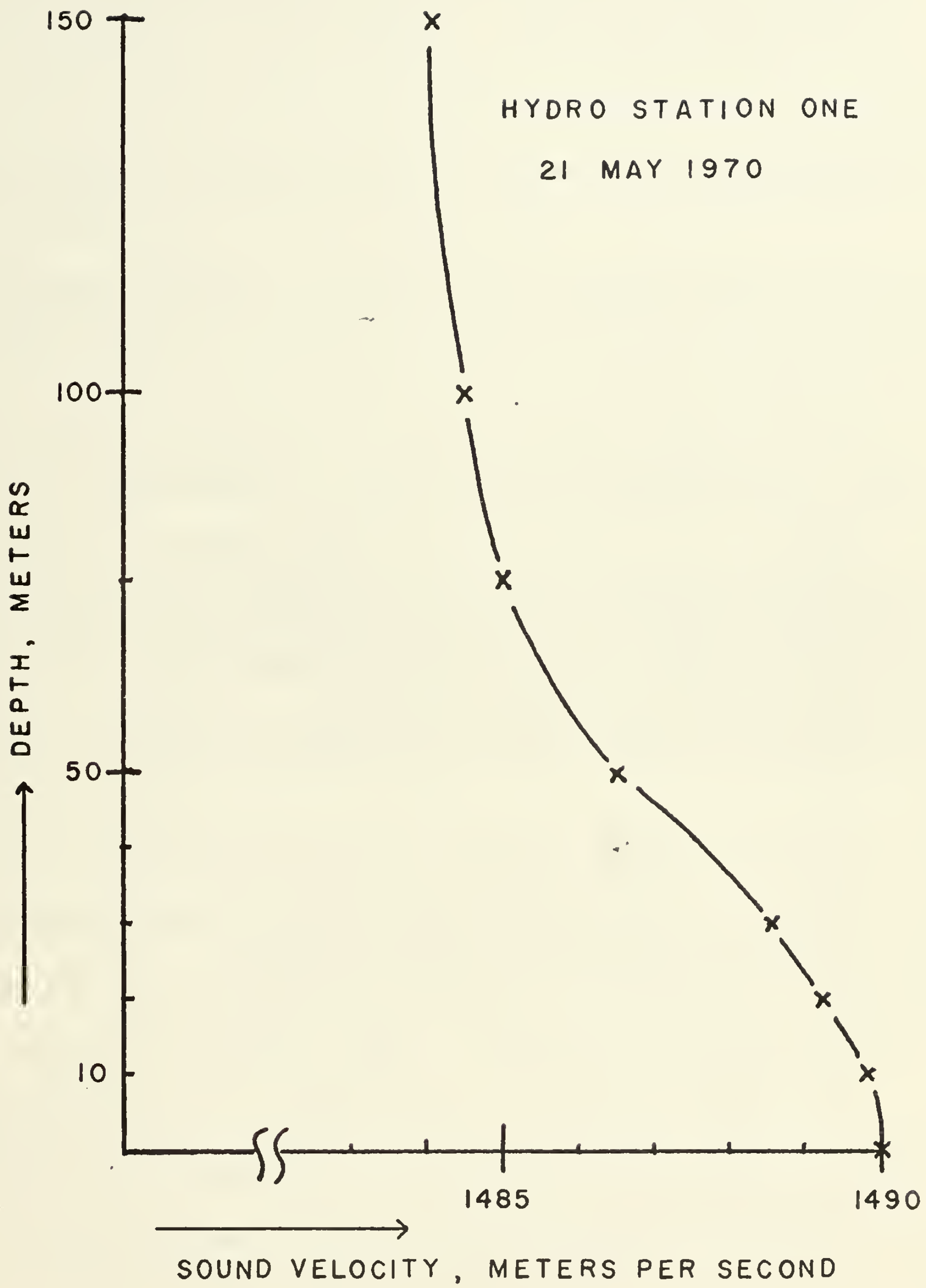


FIGURE 7

SOUND VELOCITY PROFILE

VI CONCLUSIONS

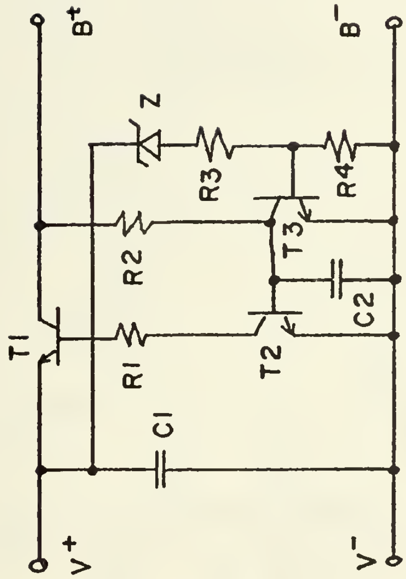
Additional development is still required to make the system as effective as theory would indicate. The multiplexing nature of the system is quite effective and makes possible the recording of numerous channels of analog data on a single-track tape recorder, but the acoustic link is not yet effective.

Additional research and development is being carried on at the Naval Medical Research Institute with the intention of re-designing the front end of the receiver. Two particular techniques are of interest; the development of a more acceptable preamplifier, and the possibility of using a type-two phase-locked loop to yield a reliable signal from an intermittent carrier. Computer work with just such a device has indicated that this may indeed be the solution to a number of problems inherent to the environment by introducing a "fly-wheel" such that when the carrier is not present the last signal received is reproduced.

The system shows potential of being an effective telemetry system capable of meeting the guidelines initially established. Its projected date for field use is August 1970, to be used by the Naval Medical Research Institute as a primary tool in obtaining and recording physiological and environmental data in support of their research program in physiological stress factors of diving.

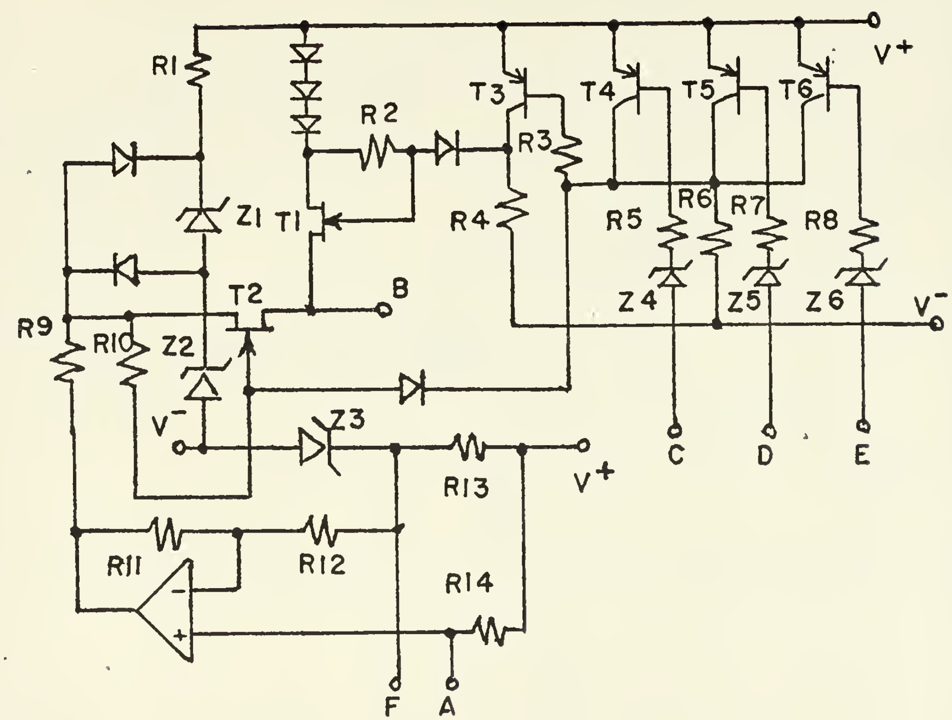
APPENDIX A

SCHEMATIC DIAGRAMS



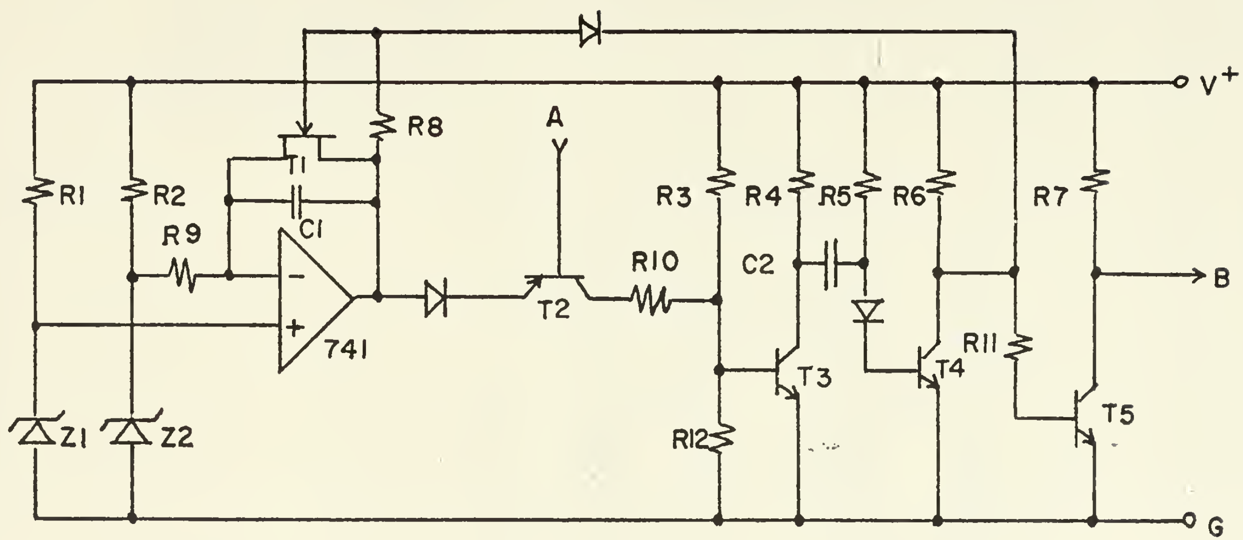
SCHEMATIC A-1

VOLTAGE REGULATOR, TRANSMITTER



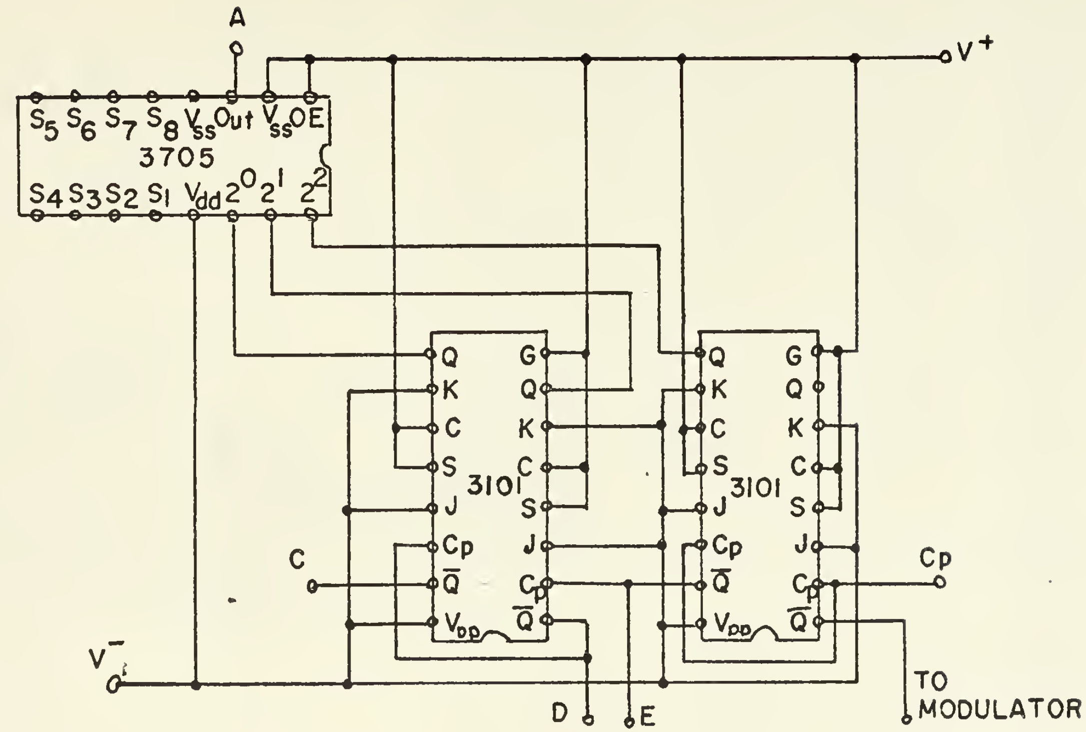
SCHEMATIC A-2

INPUT AMPLIFIER, TRANSMITTER



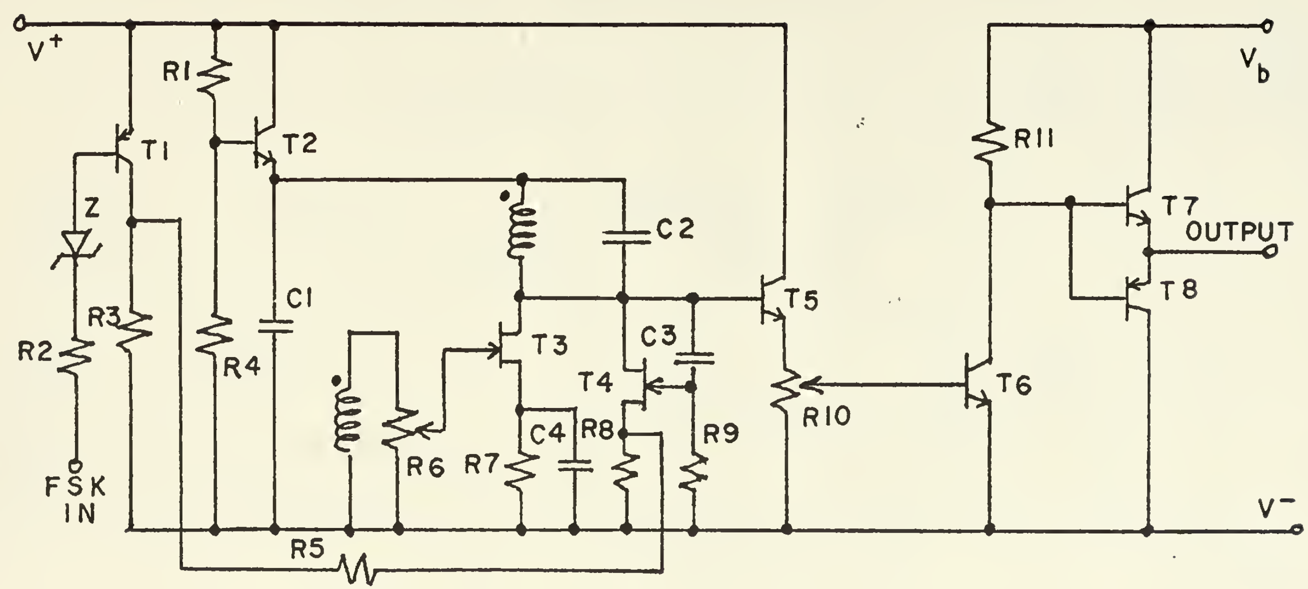
SCHEMATIC A - 3

RAMP GENERATOR AND COMPARATOR, TRANSMITTER



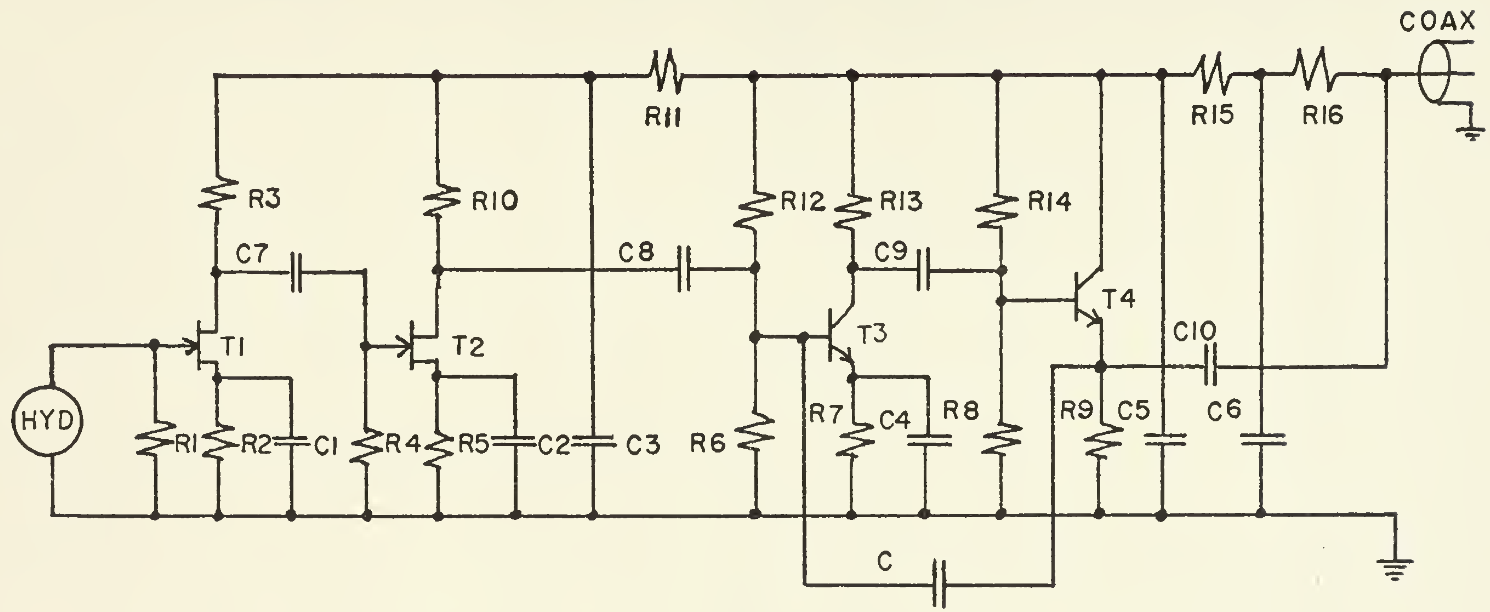
SCHEMATIC A-4

MOS INTEGRATED CIRCUITS, TRANSMITTER



SCHEMATIC A-5

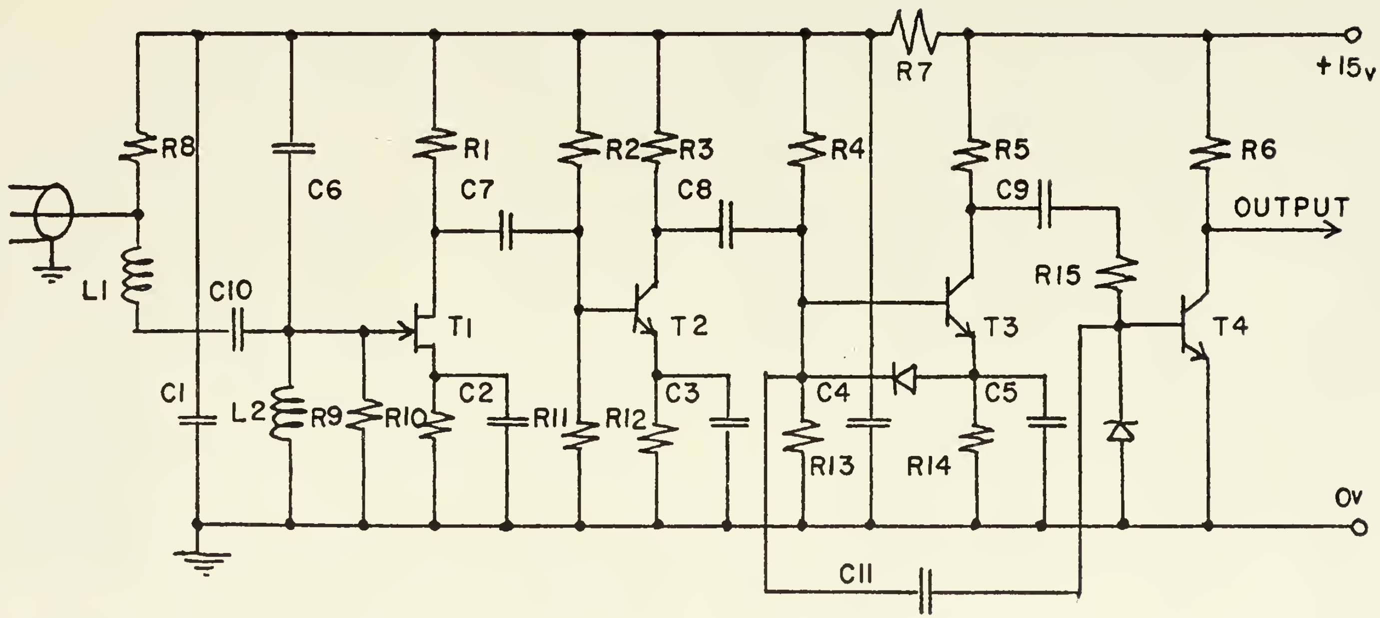
FSK OSCILLATOR, TRANSMITTER



SCHEMATIC A-6

HYDROPHONE PREAMPLIFIER, RECEIVER

R1 - 100K	C5 - 8.2uF
R2 - 3.0K	C6 - 8.2uF
R3 - 11K wire-wound	C7 - 0.0015uF
R4 - 100K	C8 - 0.0015uF
R5 - 3.0K	C9 - 0.047uF
R6 - 24K	C10 - 8.2uF
R7 - 1.6K	T1 - 2N3819
R8 - 5.1K	T2 - 2Nd819
R9 - 100	T3 - 2N3705
R10 - 11K wire-wound	T4 - 2N3705
R11 - 1.2K	C - 29pF
R12 - 100K	
R13 - 4.7K	
R14 - 51K	
R15 - 510	
R16 - 510	
C1 - 8.2uF	
C2 - 8.2uF	
C3 - 8.2uF	
C4 - 8.2uF	



SCHEMATIC A - 7

FOUR-POLE FILTER AND AMPLIFIER, RECEIVER

R1 - 7.5K

R2 -33K

R3 - 3.0K

R4 - 30K

R5 - 1.5K

R6 - 3.0K

R7 - 100

R8 - 51

R9 - 5.6K

R10 - 3.0K

R11 - 5.6K

R12 - 1.0K

R13 - 5.1K

R14 - 510

R15 - 6.2K

C1 - 8.2uF

C2 - 8.2uF

C3 - 8.2uF

C4 - 8.2uF

C5 - 8.2uf

C6 - 0.0015uf

C7 - 0.0022uf

C8 - 800pF

C9 - 0.01uF

C10 - 25pF

C11 - 1000pF

L1 - 45mH variable

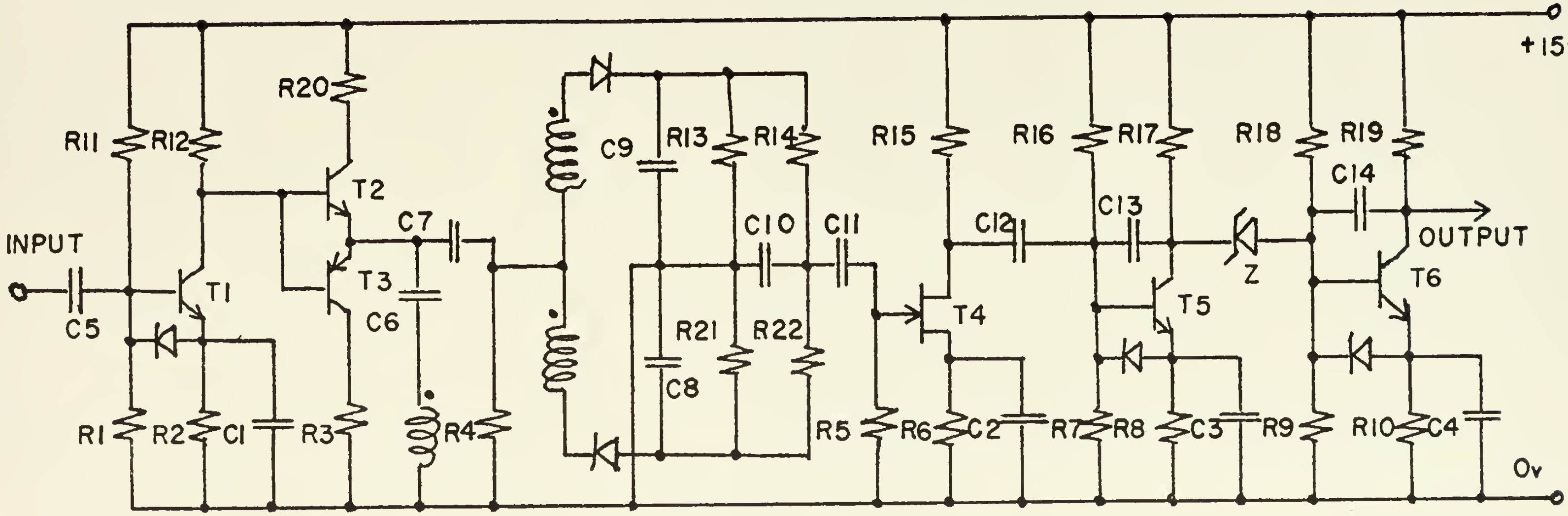
L2 - 0.85mH variable

T1 - 2N3819

T2 - 2N3705

T3 - 2N3705

T4 - 2N3705



SCHEMATIC A-8

RATIO DETECTOR AND AMPLIFIER, RECEIVER

R1 - 5.1K

R2 - 510

R3 - 100

R4 - 30K

R5 - 1.1M

R6 - 2.7K

R7 - 5.1K

R8 - 510

R9 - 5.1K

R10 - 510

R11 - 30K

R12 - 1.5K

R13 - 100K

R14 - 100K

R15 - 11K

R16 - 30K

R17 - 1.5K

R18 - 30K

R19 - 1.5K

C1 - 8.2uF

C2 - 10uF

C3 - 10uF

C4 - 10uF

C5 - 800pF

C6 - 360pF

C7 - .047uF

C8 - 800pF

C9 - 800pF

C10 - 800pF

C11 - 0.0047 uF

C12 - 10uF

C13 - 820pF

C14 - 820pF

T1 - 2N3705

T2 - 2N3705

T3 - 2N3703

T4 - 2N3819

T5 - 2N3705

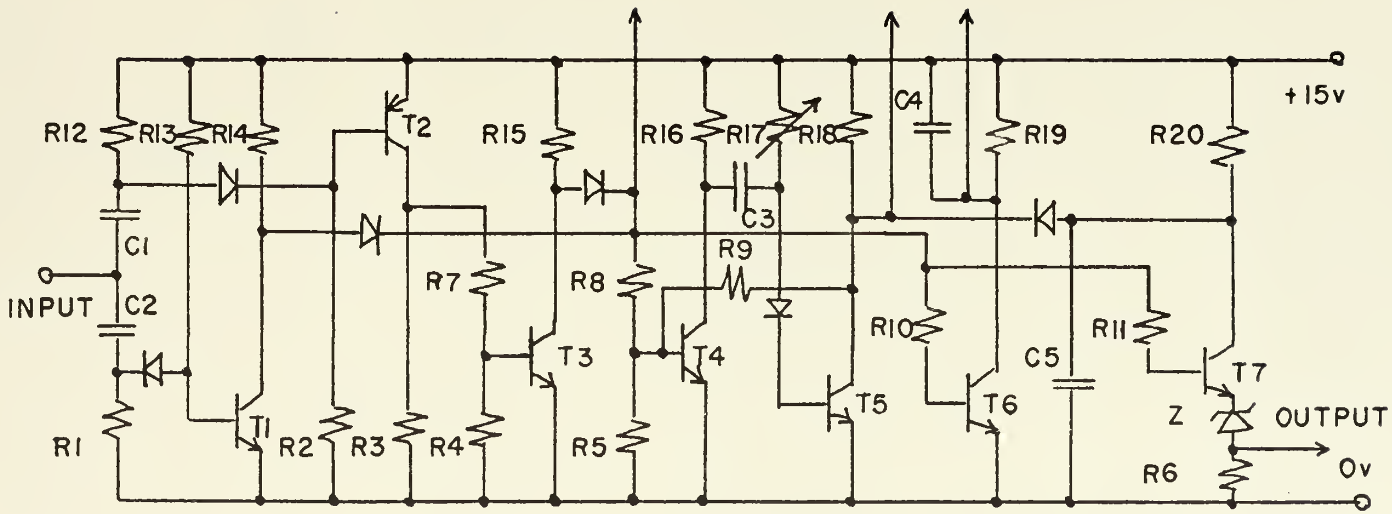
T6 - 2N3705

Z - 7-volt zener

Transformer;

$\frac{1}{2}$" O. D. ferrite toroid

3 windings, 25 turns each



SCHEMATIC A-9

DATA DISTRIBUTOR, RECEIVER

R1 - 100K

R2 - 180K

R3 - 7.5K

R4 - 100K

R5 - 100K

R6 - 100K

R7 - 51K

R8 - 7.5K

R9 - 62K

R10 - 51K

R11 - 62K

R12 - 100K

R13 - 180K

R14 - 7.5K

R15 - 7.5K

R16 - 7.5K

R17 - 50K variable

R18 - 7.5K

R20 - 51K

C1 - 0.0039uF

C2 - 0.0039uF

C3 - 0.0022uF

C4 - 0.0047uF

C5 - 0.047uF

T1 - 2N3705

T2 - 2N3703

T3 - 3N3705

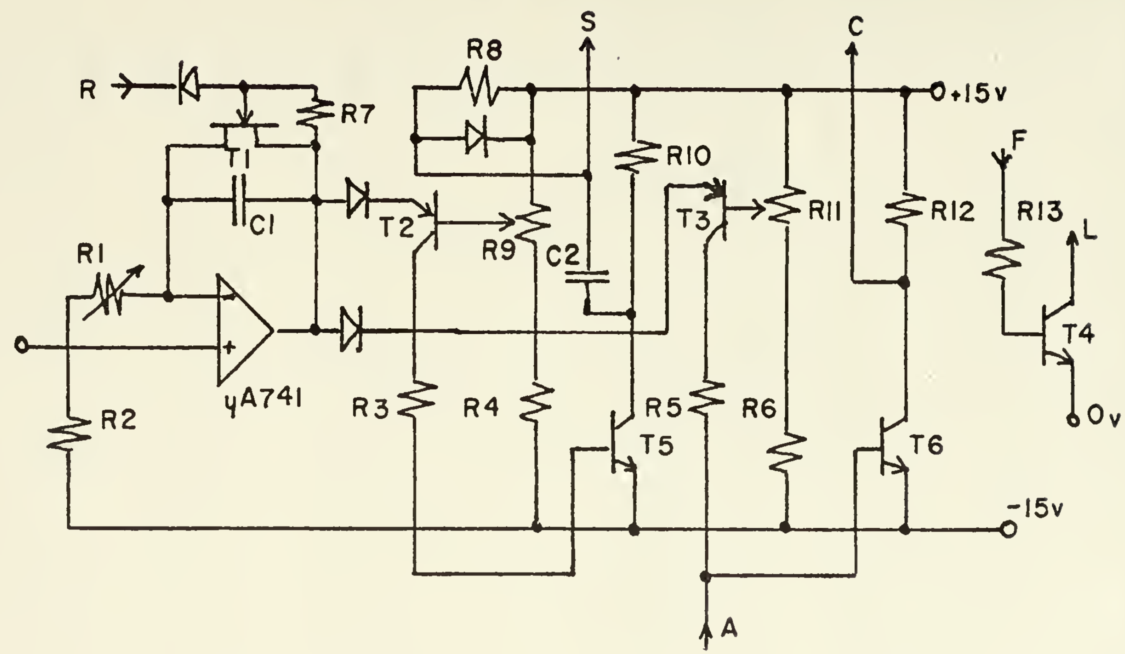
T4 - 2N3705

T5 - 2N3705

T6 - 2N3705

T7 - 2N3705

Z - 6.2-volt Zener



SCHEMATIC A-10

DATA DISTRIBUTOR RAMP GENERATOR, RECEIVER

R1 - 20K variable

R2 -30K

R3 - 30K

R4 - 30K

R5 - 20K

R6 - 30K

R7 - 1M

R8 - 100K

R9 - 20K variable

R10 - 10K

R11 - 20K variable

R12 - 10K

R13 - 20K

C1 - 0.05uF

C2 - 100pF

T1 - 2N3819

T2 - 2N3703

T3 - 2N3703

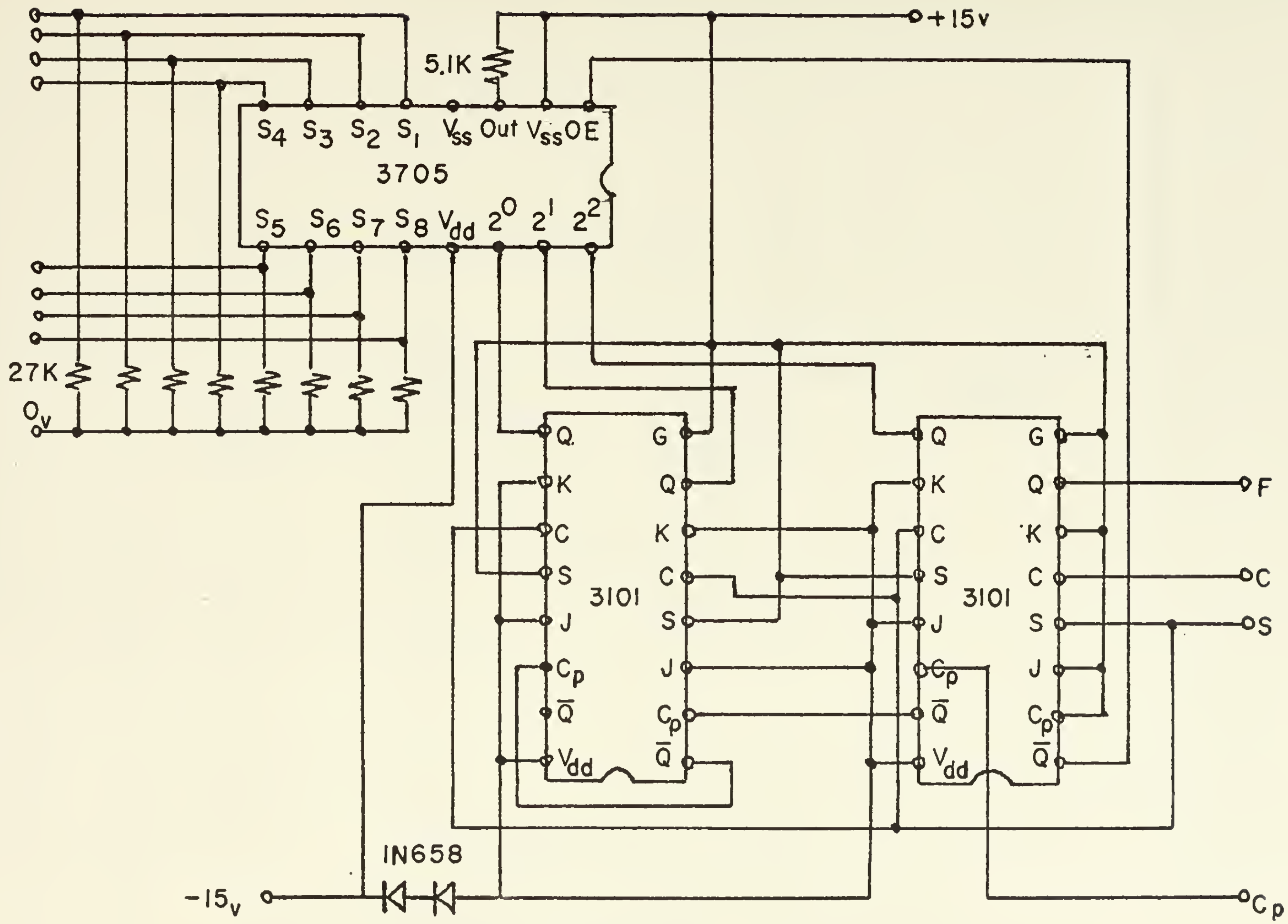
T4 - 2N3705

T5 - 2N3705

T6 - 2N3705

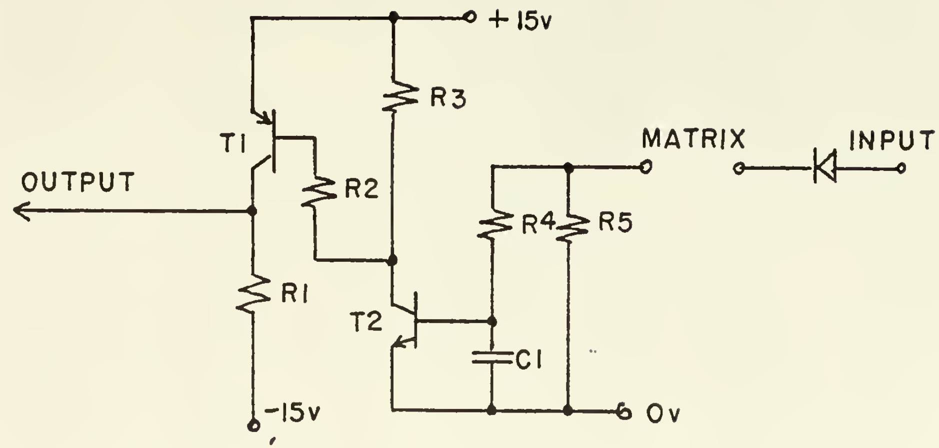
Operational amplifier;

Fairchild uA741



SCHEMATIC A-11

MOS INTEGRATED CIRCUITS, RECEIVER



SCHEMATIC A-12

SINGLE-STAGE BUFFER, RECEIVER

R1 - 7.5K

R2 - 75K

R3 - 3.9K

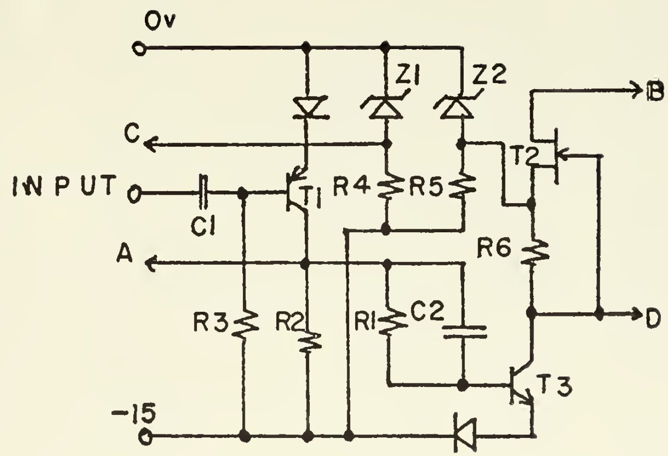
R4 - 10K

R5 - 100K

C1 - 30pF

T1 - 2N3703

T2 - 2N3705



SCHEMATIC A-13

SIGNAL CONDITIONER, RECEIVER

R1 - 51K

R2 - 10K

R3 - 240K

R4 - 6.2K

R5 - 2.0K

R6 - 51K

C1 - 0.1uF

C2 - 800pF

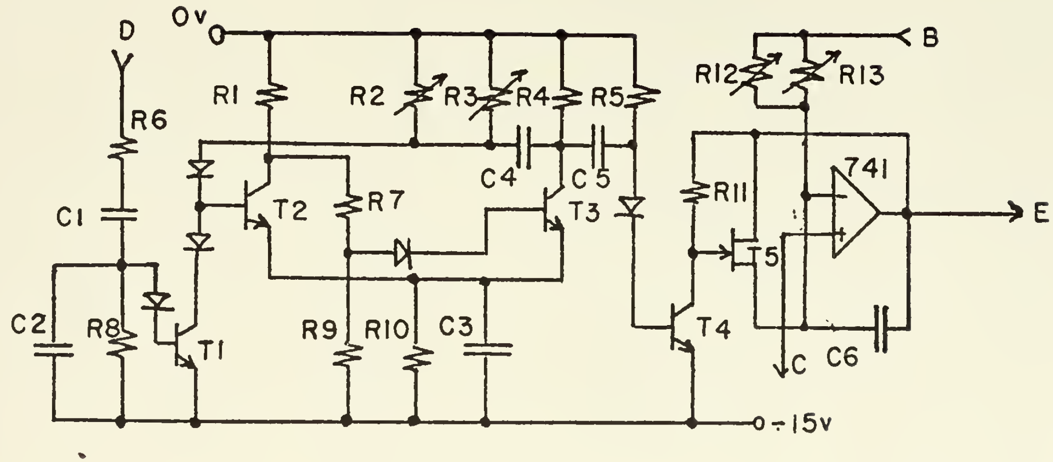
T1 - 2N3703

T2 - 2N3819

T3 - 2N3705

Z1 - 10-volt zener

Z2 - 6.2-volt zener



SCHEMATIC A-14

DEMODULATOR. RAMP GENERATOR, RECEIVER

R1 - 3.9K

R2 - 500K variable

R3 - 100K variable

R4 - 3.9K

R5 - 240K

R6 - 100K

R7 - 82K

R8 - 1M

R9 - 160K

R10 - 1.0K

R11 - 51K

R12 - 20K variable

R13 - 50K variable

C1 - 800pF

C2 - 100pF

C3 - 8.2uF

C4 - 0.03uF

C5 - 0.01uF

C6 - 0.05uF

T1 - 2N3705

T2 - 2N3705

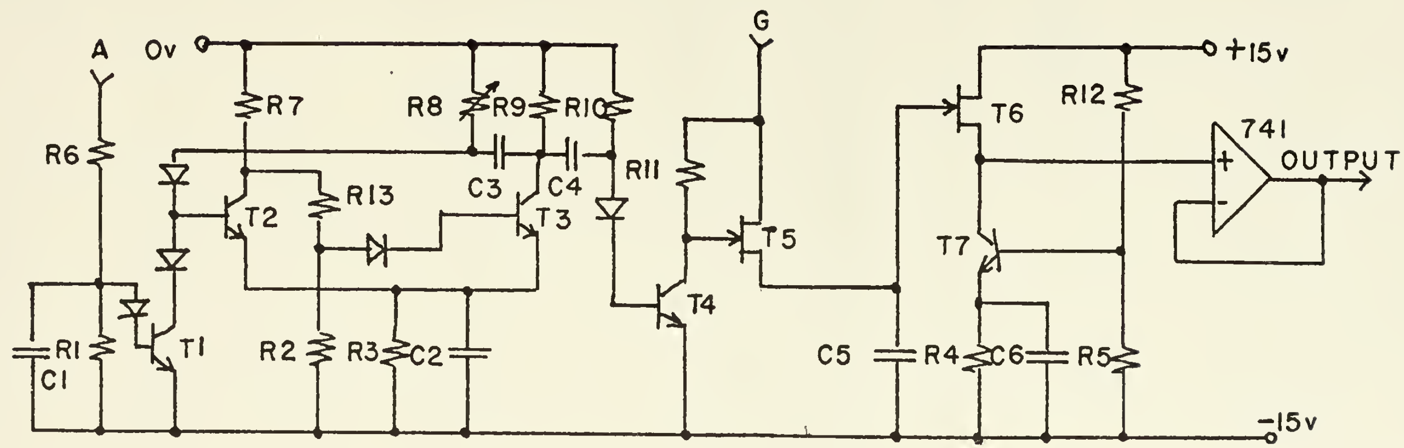
T3 - 2N3705

T4 - 2N3705

T5 - 2N3819

Operational amplifier;

Fairchild uA741



SCHEMATIC A-15

SAMPLE AND HOLD, RECEIVER

R1 - 1M

R2 - 160K

R3 - 1.0K

R4 - 1.2K

R5 - 2.4K

R6 - 100K

R7 - 3.9K

R8 - 100K variable

R9 - 3.9K

R10 - 240K

R11 - 51K

R12 - 24K

C1 - 100pF

C2 - 8.2uF

C3 - 0.0022uF

C4 - 0.0022uF

C5 - 0.1uF

C6 - 0.03uF

T1 - 2N3705

T2 - 2N3705

T3 - 2N3705

T4 - 2N3705

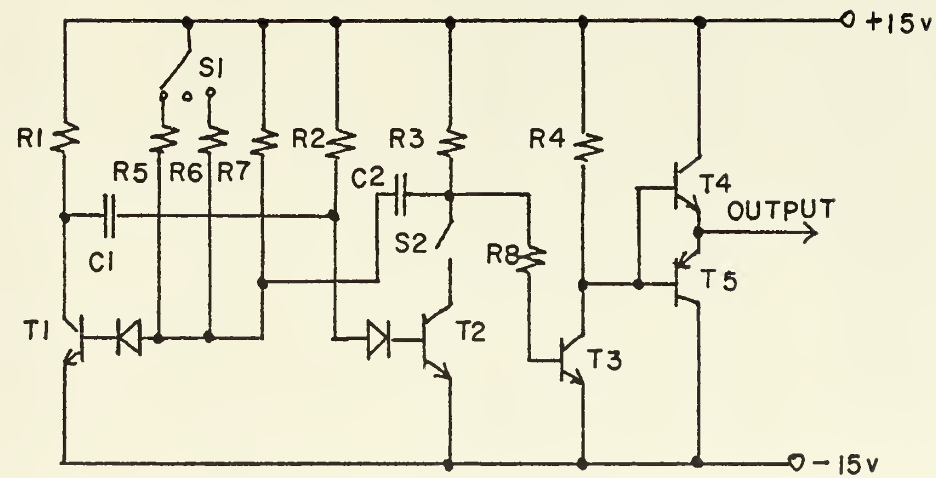
T5 - 2N3819

T6 - 2N3819

T7 - 2N3705

Operational amplifier;

Fairchild uA741



SCHEMATIC A-16

CALIBRATION UNIT, RECEIVER

R1 - 3.0K

R2 -33K

R3 - 3.0K

R4 - 3.0K

R5 - 33K

R6 - 62K

R7 - 33K

R8 - 27K

C1 - 0.068uF

C2 - 0.068uF

T1 - 2N3705

T2 - 2N3705

T3 - 2N3705

T4 - 2N3705

T5 - 2N3703

APPENDIX B.

Fairchild A741 Operational Amplifier

Input offset voltage (R_s=10 Kilohms)	1.0	5.0	millivolts
Input offset current	30	2000	nanoamperes
Input bias current	200	500	nanoamperes
Input resistance 0.3	1.0		megohms
Large-signal voltage gain (R_1=2K $V_{out}=\pm10$volts)	50,000	200,000	
Output voltage swing (R_1= 10K)	±10	±13	volts
(R_1=2K)	±12	±13	volts
Power consumption	50	85	milliwatts
Slow rate (unity gain)	0.5		volts/microsecond

Features;

A. Fully frequency compensated

B. Short-circuit protected

C. Offset voltage null capability

D. Resistant to "latch-up"

APPENDIX C.

Fairchild MOS 3705 8-Channel Analog Switch

-20v	V_{dd}	-24v
-5.0v	V_{out}	+5.0v
5.0v	V_{ss}	7.0v

Logic input "high" level

$(V_{ss}$-1.5v)	V_{ih}	V_{ss}

Logic input "low" level

V_{dd}	V_{il}	+0.2v

Power dissipation	min.	130 milliwatts
	max.	175 milliwatts
Channel "on" resistance	typ.	250 ohms
	max.	400 ohms

Features;

A. Output enable control

B. One-microsecond switching time

C. Full decoding within the device

D. Input gate protection

E. Zero offset voltage

APPENDIX D.

Fairchild MOS 3101 Dual JK Flip-Flop

$V_{dd} = -27 \pm 2$ volts

Clock amplitude	min.	-9.0 volts
Clock pulse width	min.	1.0 microseconds
Input logic "high"	V_{ih}	-9.0 volts
Input logic "low"	V_{il}	-2.0 volts
Output logic "high"	V_{oh}	-10.0 volts
Output logic "low"	V_{ol}	-1.0 volts
Power consumption	typ.	75 milliwatts
Switching time	typ.	0.5 microseconds

Features;

A. Separately clocked inputs

B. Input gate protection

C. Buffered outputs

REFERENCE

1. Campbell, D. W., Cyr, R. J., Crosier, C., "Underwater Telemetry for Oceanographic Research," Electronics, p. 53-55, 12 January 1962.

2. Thanos, S. N., Hubbard, A. C., "Two-Way Hydroacoustic Communications Link for an Ocean-Bottom Seismograph," IEEE Transactions on Geoscience Electronics, v. GE-4, p. 17-24, June 1966.

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13. ABSTRACT

A general-purpose acoustic telemetry system is presented. The carrier of one hundred forty-five kilohertz is frequency shift keyed at an average rate of once a millisecond, the subcarrier being pulse-width modulated by seven channels of analog data and one synchronization channel. The receiver is automatically synchronized and various error detecting schemes are employed.

The system is designed to be battery operated at the transmitter and is specifically intended to telemeter physiological and environmental data from a free dwimming diver to the surface. The theoretical lateral design range is three hundred meters.

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Acoustic Telemetry						
Underwater Acoustic Telemetry						
Physiological Monitoring						

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